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Fourth Bimonthly Report on the RT-21

Transmitter Development

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I Purpose

See Bimonthly Report No. 1.

II Abstract

As a result of a visit from the customer at the end of the previous reporting period, the antenna impedance specifications were modified. During the present reporting period, the changed and somewhat less stringent requirements were examined while a system was worked out on paper which would satisfy the original requirements. In this report the system meeting the full specifications is described and an account given of the relative cost, in terms of circuit and device complexity, of the compromises suggested by the reduced requirements.

To meet the original specifications, a system has been designed using a pi section for matching the transmitter to the antenna. In order to make the maximum to minimum values of the elements of the network practical, it is necessary, for most transistor load impedances, to make all three adjustable. The phase and amplitude sensing circuits supply only two control signals. The system consequently derives a control signal for the third element by means of limit switches on the other two elements, in conjunction with appropriate logic circuitry. In order to satisfy the original specifications, it is interesting to note that the resultant network is also capable of matching to a very much wider range of impedances.

A study is described which was carried out to determine whether, with a considerably simpler two-variable element pi network and practical maximum-to-minimum ratios, a sufficiently large percentage of the originally specified impedance range could be satisfied, to be acceptable to the customer. In that

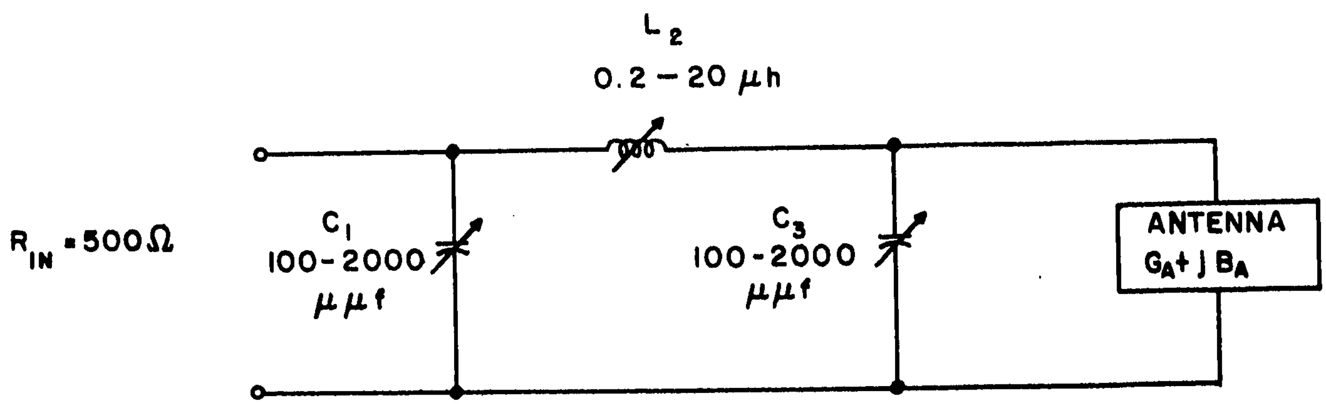
the ultimate objective of this program is to develop a transmitter and matching network capable of being packaged in a volume of 27 cubic inches, a reduction in the number of controlled elements from two to three represents a very appreciable improvement as long as the versatility of the equipment is not seriously reduced.

A system for automatic tuning of the transmitter using electrically variable elements has been developed as described in previous bimonthly reports. During the past reporting period the electronic bandswitching circuitry was constructed and coupled to the voltage variable capacitor control system. The results of this work are described in this report as well as a description of the modifications which were found to be necessary in the circuitry described previously. Now that an operative, but quite complex, system has been constructed, it will be studied critically to determine ways in which simplifications can be made without sacrificing performance capabilities.

III Factual Data

1. Three-Variable Pi Network

It has been shown in previous reports that, if a pi network having two-variable elements is used to match the antenna impedance rectangle over the 3-30 mc band, variable elements with impractical ranges are required. However, if all three elements in the pi could be varied, the impedance matching can be accomplished with elements which are realizable. An analysis of the circuit equations demonstrates that the network shown in Figure 1 will match the specified impedance rectangle to 500 ohms over the 3-30 mc band. Elements having this range are presently available, but their large physical size makes them undesirable. However, it appears that special components of acceptable size



THREE-VARIABLE IMPEDANCE MATCHING NETWORK

FIGURE 1

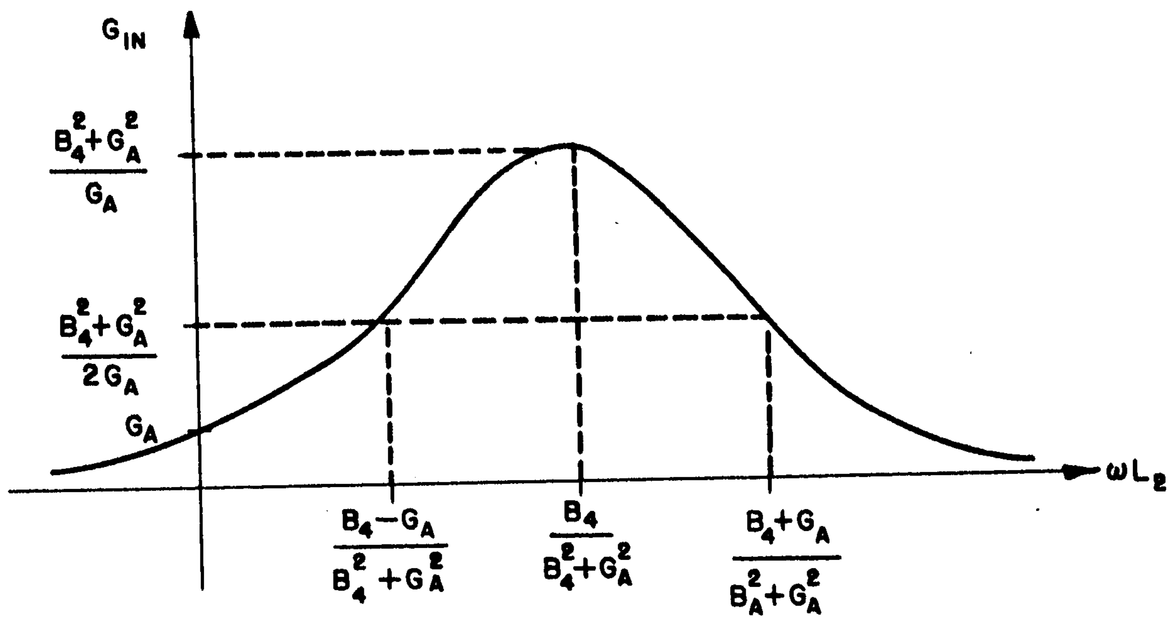
could be developed.

Assuming that the required components can be obtained, it is necessary to devise a method of controlling the three variable elements. The impedance detector provides two signals; one indicates impedance phase and the other indicates impedance magnitude. These two signals can be used to control two of the variable elements in the pi. The third variable element can be controlled by employing switches which are operated whenever either of the two detector controlled variables is driven to its maximum or minimum position. These switches will then activate the servo of the third variable until the two detector controlled variables have left their extreme positions. This obviously releases the limit switches, and the third servo ceases operation.

It appears that the system will operate most satisfactorily if the phase detector is used to operate C_1 and the magnitude detector is used to operate L_2. The operation of the servo system can be determined by examining the manner in which input conductance and susceptance vary as a function of L_2. The conductance curve is shown in Figure 2, and the curve showing the value of ωC_1 necessary to produce zero phase ~~single~~ ANGLE is shown in Figure 3. On these curves B_4 is the sum of $\omega C_3 + B_A$. In order to be sure that the solution $Y_{in} = G_{in} + j\,0 = G_0$ exists, certain constraints must be imposed. The first constraint is that the inequality

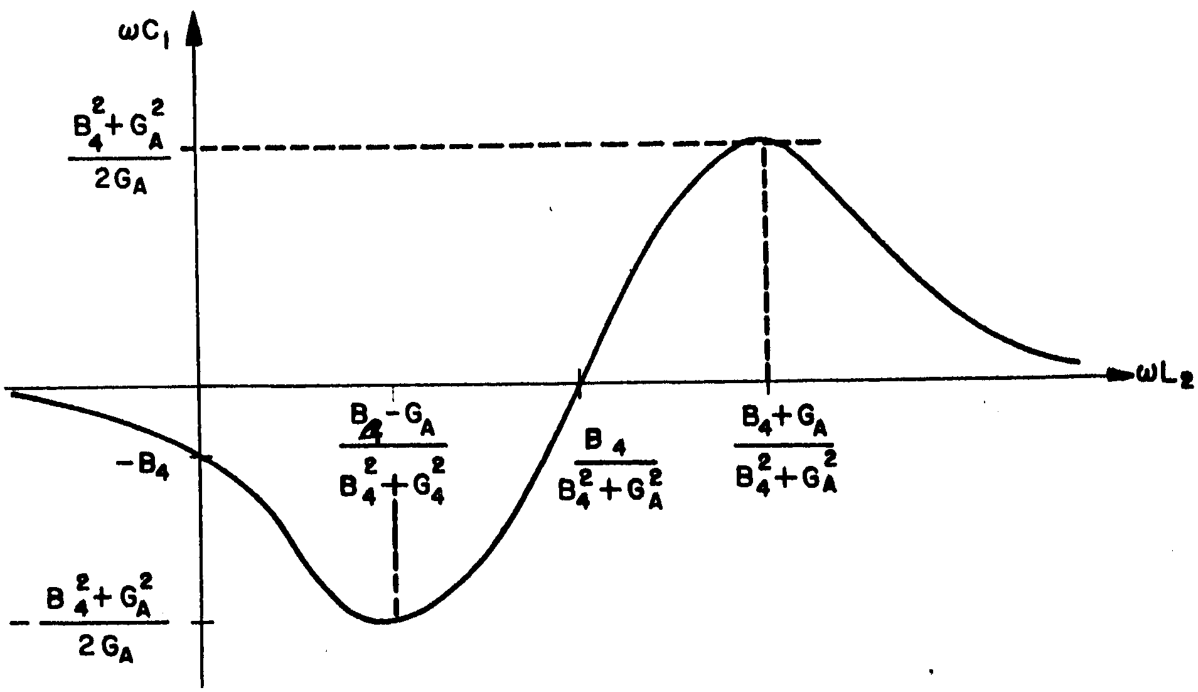
$$\frac{B_4^{\,2} + G_A^{\,2}}{G_A} \geq G_0$$

must be satisfied if the point $G_{in} = G_0$ is to lie on the conductance curve. This can be insured by prepositioning C_3 to its maximum value. The second constraint



INPUT CONDUCTANCE AS A FUNCTION OF ωL_2

FIGURE 2



C_1 SUSCEPTANCE REQUIRED TO PRODUCE ZERO PHASE ANGLE AS A FUNCTION OF ωL_2

FIGURE 3

arises from the fact that there are two points on the conductance curve where $G_{in} = G_0$. However, for the point nearest the G_{in} axis, a negative value for C_1 may be required. Therefore the servo system should be connected so as to select the solution where ωC_1 is always positive. Operation on the desired portion of the conductance curve can be insured by prepositioning L_2 to its maximum position. This also makes $G_{in} < G_o$ when the servo system begins operation. The magnitude detector should then be connected in a manner which causes L_2 to decrease if $|Y|_{in} < G_0$ and increase if $|Y|_{in} > G_0$. Similarly, the phase detector should cause C_1 to decrease if the phase is capacitive and increase if the phase is inductive.

The desired effect of the limit switches of C_1 and L_2 may be determined by assuming the system is ready to start operation. As a result of prepositioning, ωL_2 and B_4 are initially at their maximum. C_1 may initially be either tuned or driven minimum. It is certain that initially $G_{in} < G_0$, but if C_1 has been driven minimum, it may be that $|Y_{in}| > G_0$. The magnitude detector would try to increase L_2 in this event, but this is opposite to the desired effect. This difficulty can be overcome by having the C_1 minimum limit switch remove the control of L_2 from the magnitude detector and instead apply a voltage which causes L_2 to decrease until C_1 leaves its minimum position. One potentially unfavorable situation should be investigated at this point. If $G_{in} > G_0$ when the C_1 minimum limit switch returns the control of L_2 to the magnitude detector, L_2 will increase and drive C_1 right back to its minimum position. This cycle would keep repeating indefinitely and no solution would ever be reached. However, it can be shown that this situation cannot occur, because the value of ωL_2 at which C_1 leaves its minimum is always greater than the value of

ωL_2 at which G_{in} reaches G_0. A sufficient condition to insure this is that

$$\omega\left[C_1\right]_{min} \leq 10\,G_0$$

Now as L_2 continues to decrease, G_{in} approaches G_0. However, as L_2 decreases, the required value of C_1 necessary to produce zero phase increases. Eventually C_1 may be driven to its maximum before G_{in} reaches G_0. Since the required value of capacitance can be reduced by decreasing C_3, the switch which is operated when C_1 reaches its maximum should be connected in a manner which accomplishes this. C_3 can always be decreased enough to cause C_1 to leave its maximum. One other possibility is that L_2 will be reduced to its minimum without causing G_{in} to reach G_0. This situation can also be corrected by reducing C_3, so that the limit switch which is operated when L_2 reaches its minimum should be connected in a manner which accomplishes this. C_3 stops decreasing when both C_1 and L_2 have left their maximum and minimum position, respectively. At this point G_{in} has reached G_0, the phase angle has become zero, and the system comes to rest.

As a summary, the system operation may be outlined as follows:

(i) L_2 and C_3 are electrically prepositioned to their maximum values.

(ii) The variables in the pi then move as dictated by these commands:

(a) The phase detector causes C_1 to decrease if the phase is capacitive and to increase if the phase is inductive. The switch at C_1 minimum causes L_2 to decrease, while the switch at C_1 maximum causes C_3 to decrease.

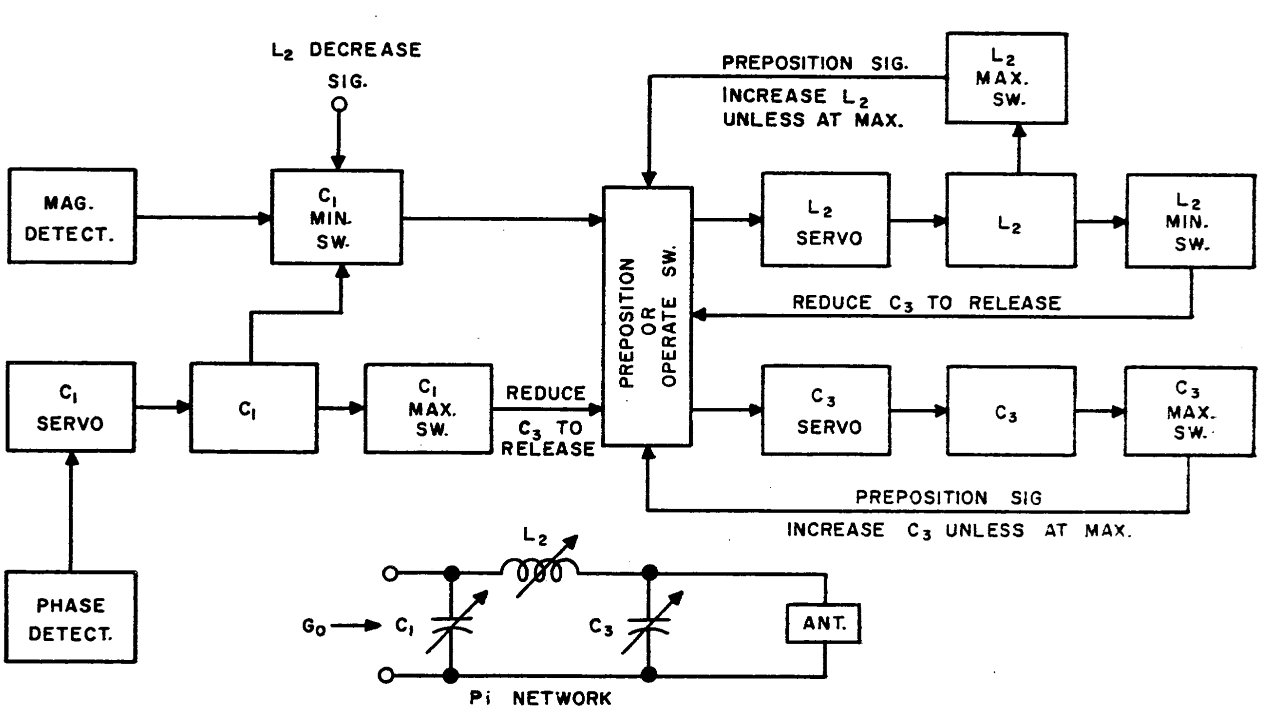
(b) The magnitude detector causes L_2 to decrease if $|Y|_{in} < G_0$; if $|Y| > G_0$, the magnitude detector causes L_2 to increase provided C_1 is not at its minimum. The switch at L_2 minimum causes C_3 to decrease.

This system is shown in block diagram form in Figure 4.

The three-variable pi appears to represent a solution to the problem of matching the specified impedance rectangle over the entire frequency band. One difficulty is that the required components must be "custom built" if reasonable physical size is to be achieved. Since the tuning ranges are functions of the desired input resistance of the network, it is questionable whether such a system should be built before the output impedance of the transistor is definitely known. The values stated were the result of assuming a 500 ohm input. As the desired input resistance is reduced, the range required of the capacitors is made more severe. It appears that a capacitor having a range of 100 - 4500 $\mu\mu$f would be required for an input resistance of 50 ohms.

It is also anticipated that considerable care must be exercised in the design of the switching circuitry. If small servo motors are to be used, a very limited amount of torque is available. Thus, not only must the limit switches operate at precisely the right time, but the operation must require only a small amount of torque.

One feature of the impedance matching capability of the pi is worth noting. In the course of evaluating the limits of required component sizes, the points of main concern were where G_A was either at maximum or minimum. These extreme values would have been unchanged had the admittance area (as shown in Figure 5) been modified by adding the area introduced by regarding the dotted line at $G_A = 25$ as part of the boundary. However, the mapping of this line into the impedance plane shows that a much larger area than was originally specified can be matched. This increased capacity is shown in Figures 6 and 7.



BLOCK DIAGRAM OF THREE—VARIABLE PI IMPEDANCE MATCHING SYSTEMS

FIGURE 4

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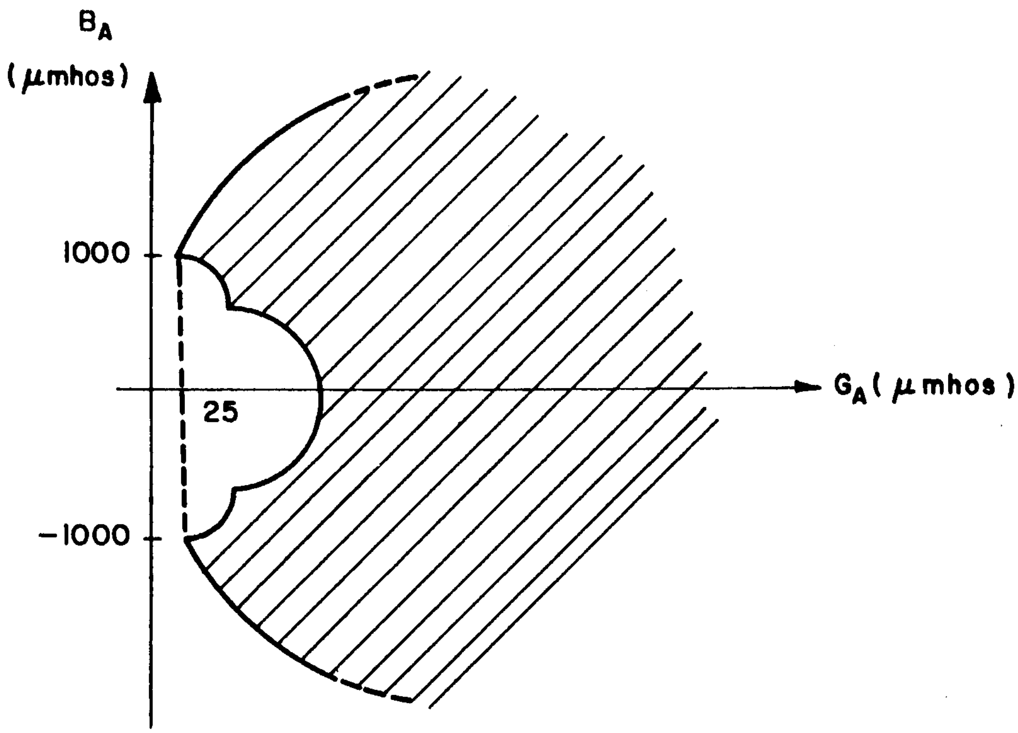
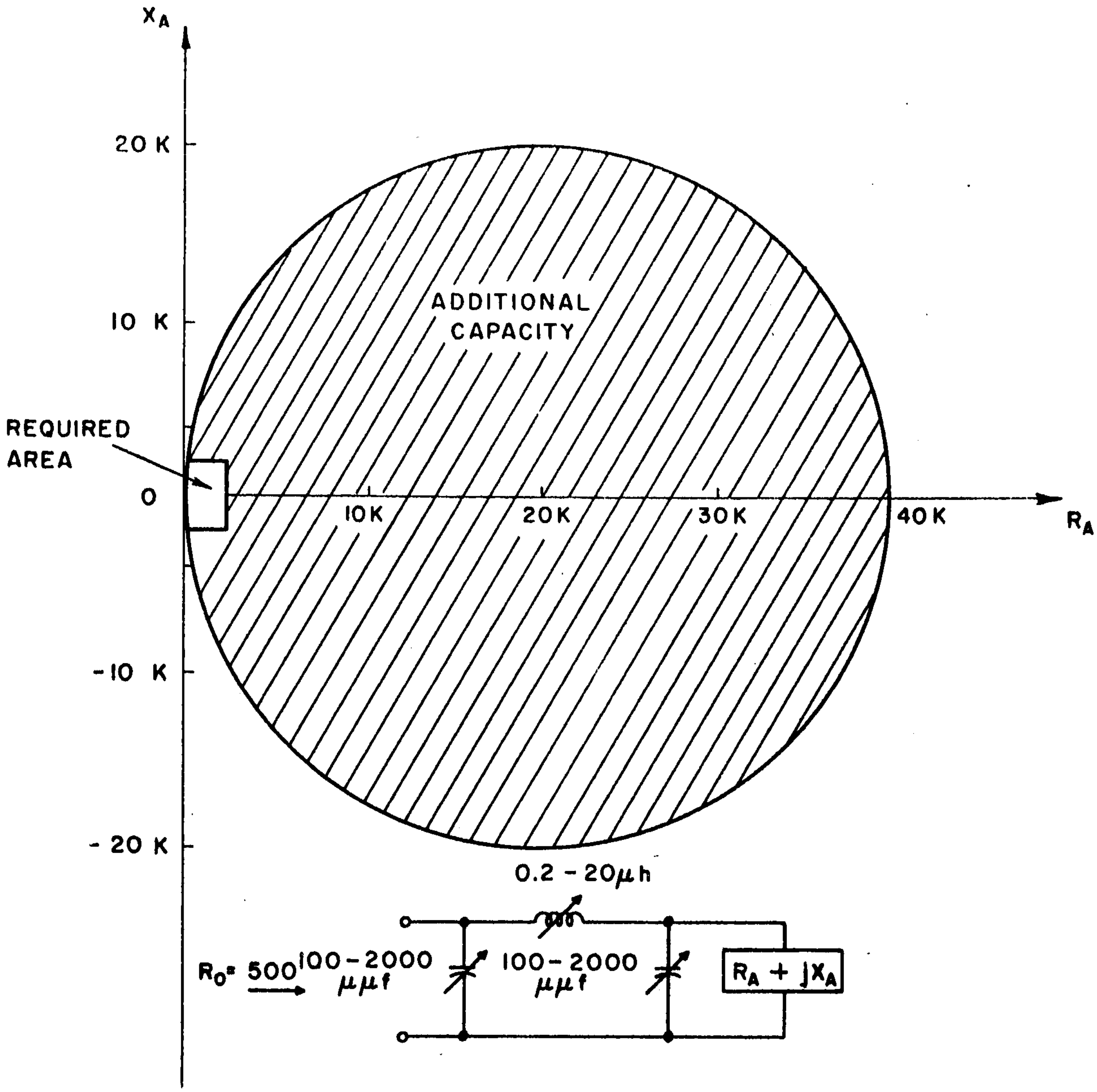


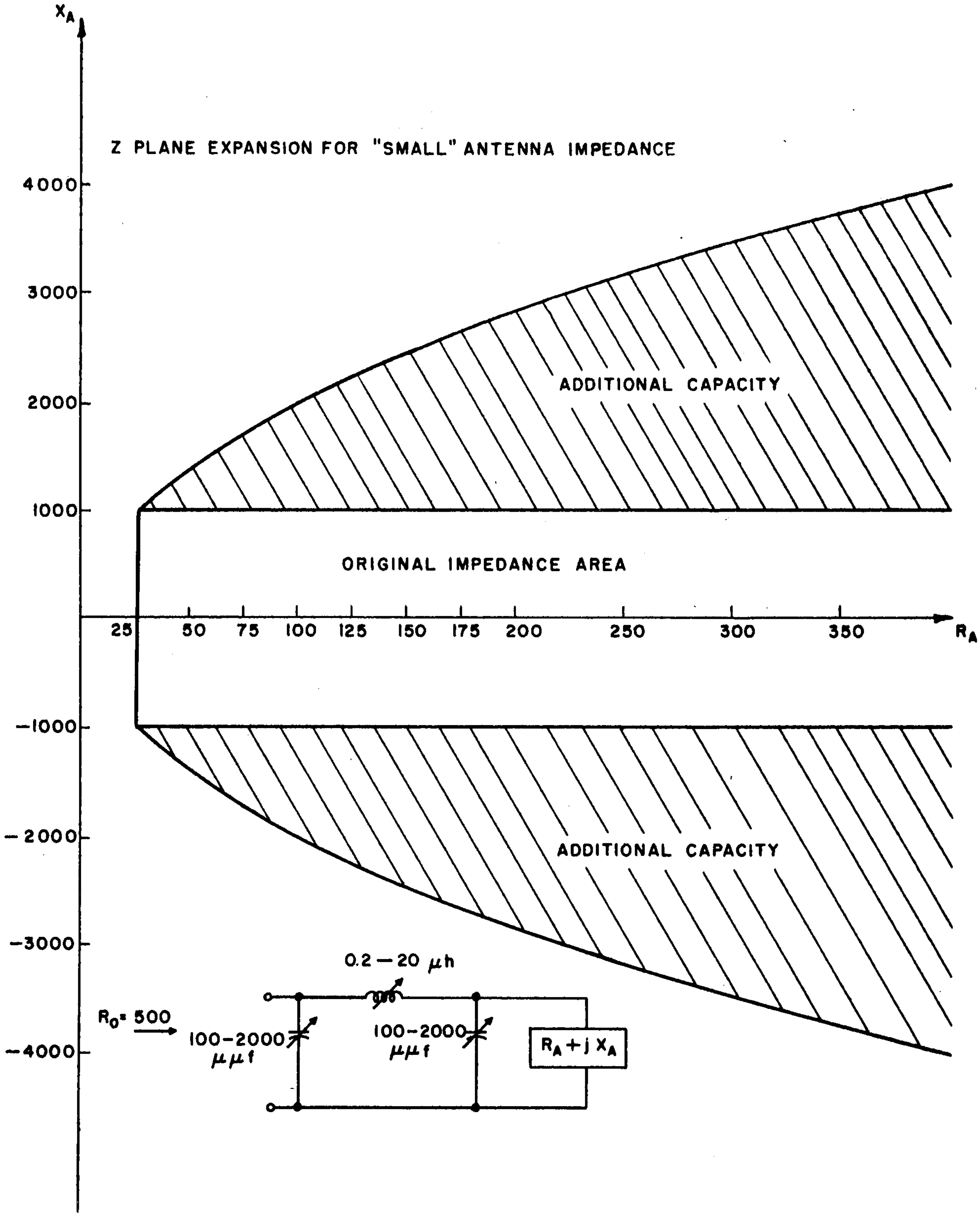
ANTENNA ADMITTANCE NEAR ORIGIN

FIGURE 5



3-30 MC IMPEDANCE MATCHING CAPABILITIES OF π

FIGURE 6



3—30 MC IMPEDANCE MATCHING CAPABILITIES OF π

FIGURE 7

2. Two-Variable Pi Network

The new specifications for the antenna impedance were interpreted to mean that the impedance is a function of frequency. This function for both the 35 foot slant wire and 25 x 25 foot inverted "L" is a spiral in the R-X plane. Under the circumstances, it is possible to design a two-variable pi network that has an input resistance of 500 ohms to 1000 ohms over the frequency range of 3 to 30 mc. The variable components of the pi are:

L - .39 to 2.7 μh. C_1 - 47 to 268 $\mu\mu$f.	for 15 to 30 mc C_2 = 100 $\mu\mu$f.
L - .50 to 8.6 μh. C_1 - 158 to 2080 $\mu\mu$f.	for 3 to 15 mc C_2 = 300 $\mu\mu$f.

This network is shown in Figure 8. It may be seen that the above network will also satisfy the degenerate case where $X_A = 0$ and $R_A = \phi(\omega)$; that is, the antenna resistance varies with frequency and the antenna reactance remains constant at $X_A = 0$.

If a similar network is used to match a rectangular area in the R-X plane bounded by R_A = 25 to 1300 ohm and $X_A = \pm$ j 1000, the values of the individual elements are a function of the input resistance of the network. The values of the variable elements are tabulated below for values of X_A between -j 1000 and +j 1000.

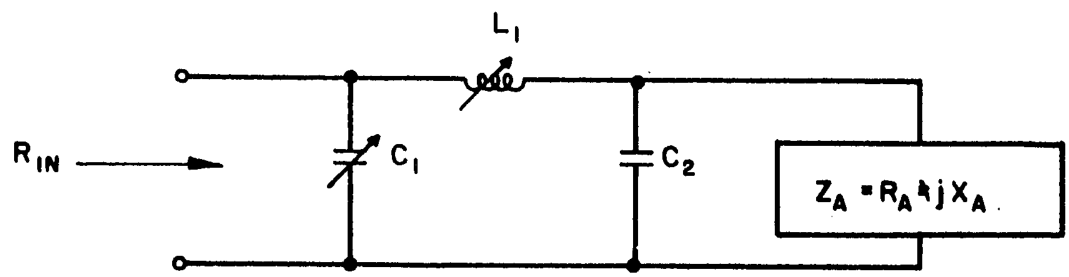


FIGURE 8

R_{in}	R_A	C_2	$L\mu h$		$C_1 \mu\mu f$		f
ohm	ohm	$\mu\mu f$	min	max	min	max	mc
50	25	1500	1.87	4.27	438	43898	3
50	25	1500	.07	.18	1043	42713	15
150	25	900	3.1	10.2	164	15562	3
150	25	900	.43	.43	361	14880	15
500	25	400	7	30.	24.5	4051	3
500	25	400	.304	1.48	89.9	3673	15
1000	25	300	2.4	67.3	36.9	2232	3
1000	25	300	.39	2.6	50	1965	15
2000	25	300	9.7	92.6	37.2	1579	3
2000	25	300	.44	3.47	35.9	1389	15

In a recent conference with the customer, it was brought out that the original interpretation of the spiral was erroneous in that the spiral was meant to be a boundary for the low resistance side of the rectangle. This would lead to a modified rectangle shown in Figure 9. This figure was made by superimposing the worst case of the spiral for values of X_A between +j 1000 and -j 1000 upon the rectangle and assuming that the spiral is no longer a function of frequency. Under these circumstances a network can be designed, the values of the individual elements again being functions of the input resistance. The circuit is shown in Figure 8 while the values for the elements are tabulated below for the 3-15 mc range.

R_{in}	$L\mu h$		$C_1 \mu\mu f$		C_2
ohm	min	max	min	max	$\mu\mu f$
50	.082	3.74	1081	14304	1500
250	.19	12.8	230	5161	700
500	.36	27.4	124	1326	400
1000	.52	48.5	51.4	732	300
2000	.61	68.4	45.0	518	300

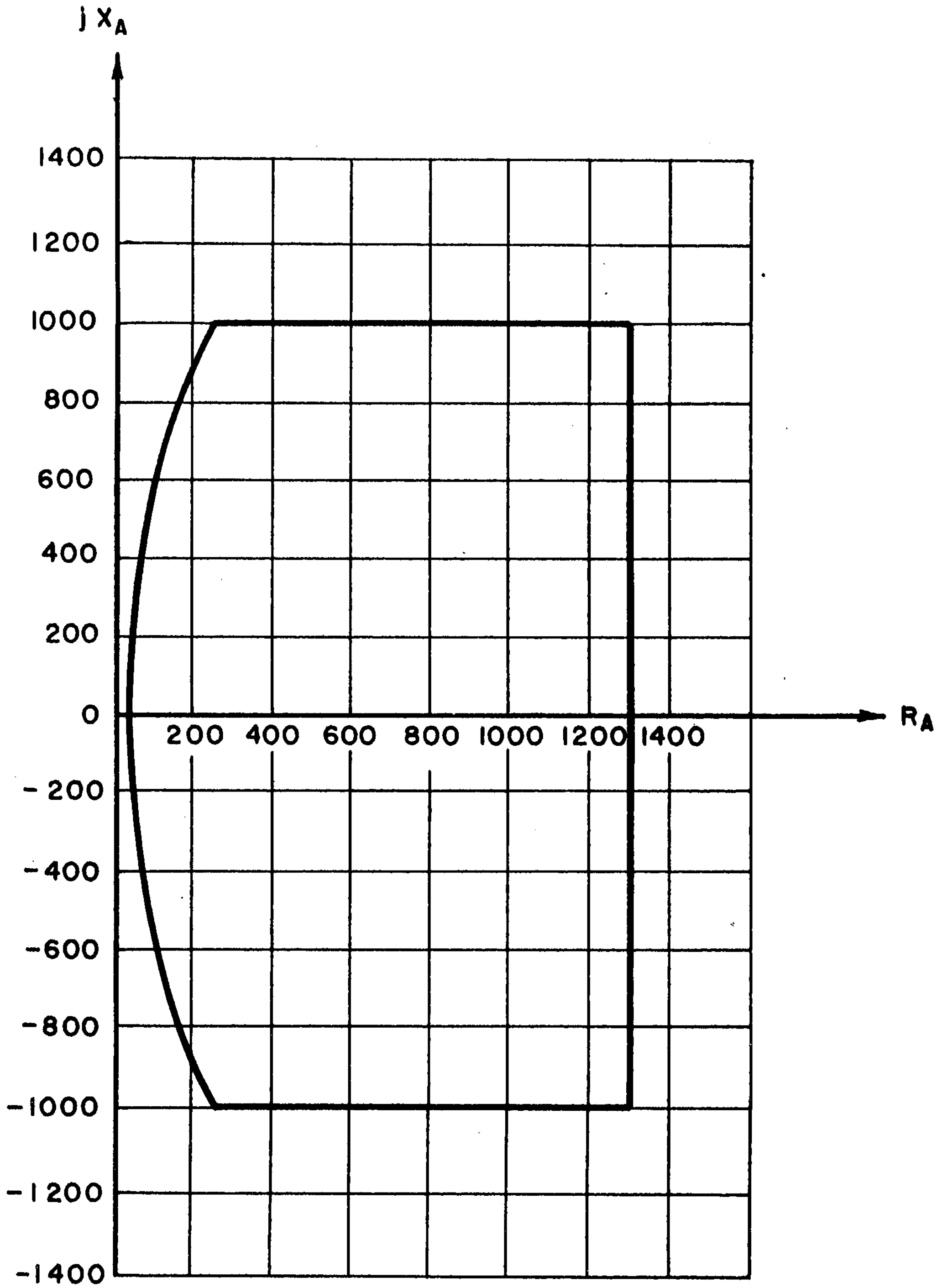

j X_A
1400
1200
1000
800
600
400
200
0
- 200
- 400
-600
-800
-1000
-1200
-1400
200 400 600 800 1000 1200 1400
R_A

FIGURE 9

This information clearly indicates that a high output impedance power amplifier would be desirable from the point of view of physical realization of the network. It will be noted that for a ten watt output the capacitors, using presently known types, must be mechanically tunable in order to accommodate the large voltage swing.

An antenna impedance of 25 +j 1000 ohms is equivalent to a parallel R-L combination in which the R is 40,000 ohms. The voltage across the parallel combination and network capacitors is:

$$E = \sqrt{WR} = \sqrt{(10)\ (40000)} = 634 \text{ v. r.m.s.}$$

As no electronically controlled capacitors are available that can handle such large voltages, a mechanical capacitor must be used. Furthermore, a minimum to maximum ratio of capacitance of about 20 and a maximum capacitance of 2000 $\mu\mu$f. will be required. This capacitor will have to be specially designed since no capacitor with this range of values is commercially available in miniature form. It is anticipated that a size eight, 400 cycle, servo motor and amplifier will be sufficient to drive a gear train linked to this variable capacitor. A preliminary servo analysis has been made that indicates that the variable elements of the pi network can be positioned sufficiently accurately with an amplifier with a gain of 300 and gear reduction ratio of 200:1. Further work in this area is pending the decision of the customer concerning which set of antenna impedance specifications it is deemed are most desirable.

3. Automatic Transmitter Tuning

In the "Third Bimonthly Report", page 20, a brief description was given of how

(i) $V_{control}$ was made to cycle between .2 volts and 25 volts.

(ii) the tank-circuit inductance L was made to follow the three-step cycle L_1, L_2, L_3, L_1, L_2, etc.

The two operations are performed with the circuitry shown in Figure 10. The capital letters in the following text refer to the blocks in the Figure 10.

(i) Cycling

Block A shows the capacitor charged by the rectified output of the balanced modulator. The voltage across the capacitor ($V_{control}$) varies from -23.8V to +1V. When $V_{control}$ approaches ground-voltage, the diode in block B will start to conduct. The resistors in block B are chosen such that when $V_{control}$ = +1V, the transistor turns off. As the collector voltage approaches -22.5V, during the short time that the transistor is turned off, the transistors in block C and block I conduct.

When conducting, the transistor in block C shorts out the 220K-resistor. Its emitter voltage drops from -1V to about -13V thus changing state of the flip-flop made up of blocks D and E. The collector voltage swing of the D-transistor is from -1.5V to -9V so that the diode (between C and D) only is conducting during the short time that the triggering lasts (and $V_{control}$ is about +1 Volt).

The flip-flop is bistable. When the D-transistor is conducting, the transistor in block H is kept turned on. The connection from block H to the balanced modulator is pulled from -16V up to ground, and the A-capacitor will discharge.

When $V_{control}$ has dropped to -23.8V the transistor in block G is turned off and its collector-voltage swings from about -22V almost up to ground-voltage. The voltage swing (which lasts as long as $V_{control}$ is around -23.8 Volts) makes the transistors in blocks F and K conduct.

Via block F the flip-flop is triggered so that the E-transistor is turned on (and the D-transistor turned off). Consequently the H-transistor is turned off

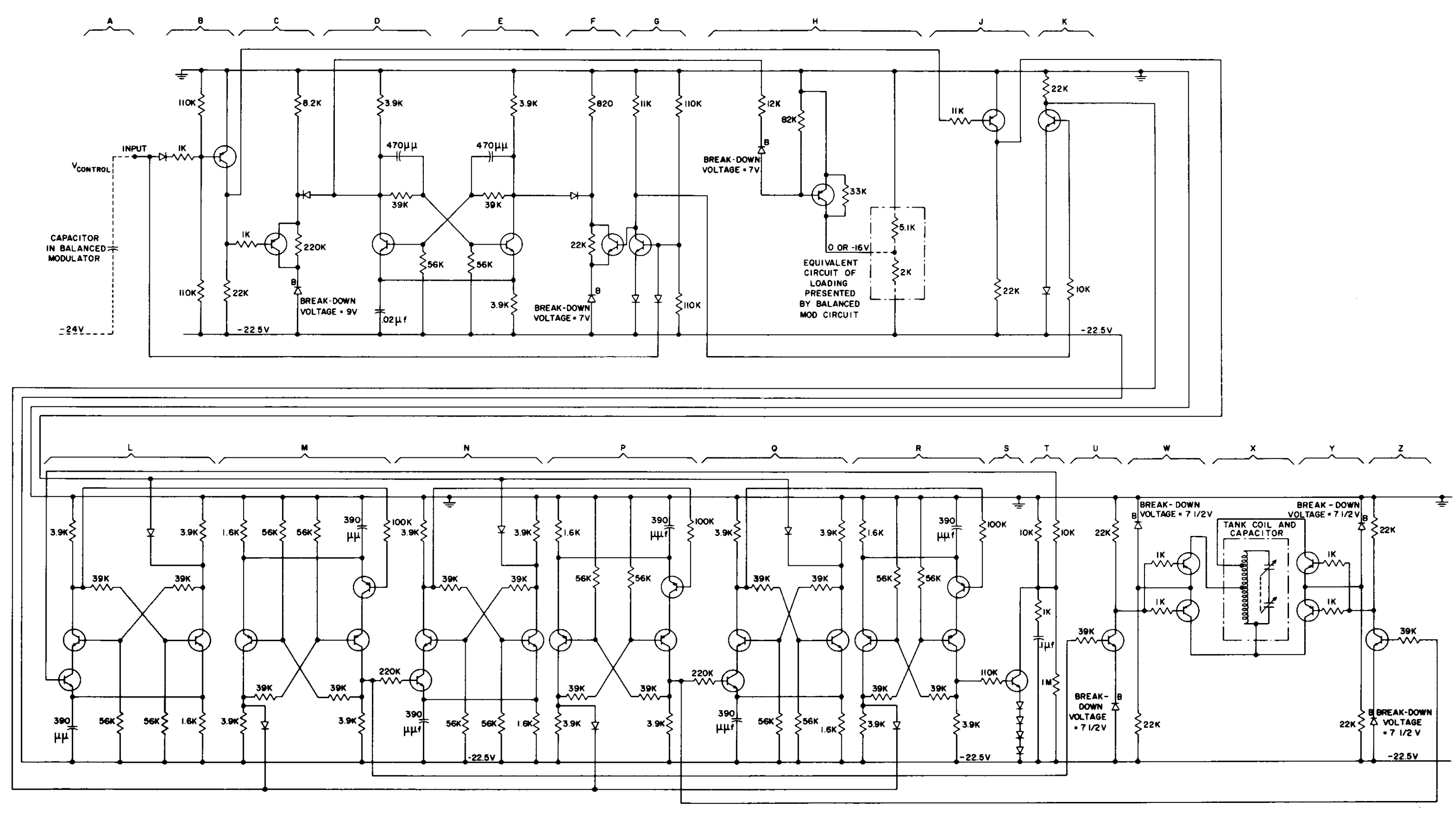

A
B
C
D
E
F
G
H
J
K
INPUT
$V_{CONTROL}$
CAPACITOR IN BALANCED MODULATOR
-24V
-22.5V
BREAK-DOWN VOLTAGE = 9V
BREAK-DOWN VOLTAGE = 7V
EQUIVALENT CIRCUIT OF LOADING PRESENTED BY BALANCED MOD CIRCUIT
0 OR -16V
L
M
N
P
Q
R
S
T
U
W
X
Y
Z
TANK COIL AND CAPACITOR
BREAK-DOWN VOLTAGE = 7 1/2V

FIGURE 10

(and stays turned off) and the balanced modulator starts to charge the A-capacitor.

When $V_{control}$ has reached +1 Volt, the B-transistor is turned off etc., as described before.

$V_{control}$ will thus continue to cycle between -23.8 Volts and +1 Volt.

The flip-flop D and E is the memory which keeps the voltage of the connection from block H to the balanced modulator at ground-voltage or at -16V. Referring to Figure 3 of the "First Bimonthly Report", this is accomplished by the switching action of H in the return leg of the balanced modulator. When H collector is at -16V the modulator passes the full 2.2 kc signal to be rectified and charge capacitor A. When the H collector is at ground the modulator balances and no output is left to charge the A capacitor, which then begins to discharge through its associated time constant.

The lower transistor is the balancing switch. Transistor H in Figure 9 is effectively in parallel with this switch and can be made to balance the balanced modulator independently. This action can interrupt the output from the 2.2 kc oscillator through the balanced modulator and terminate charging of capacitor A as soon as block H transistor collector is at ground potential. When the H collector is at -16V it is calling for charge on capacitor A, whereas when it is at ground it is calling for capacitor A to discharge.

<u>(ii) Tank Circuit Switching</u>

It will be noticed that each time $V_{control}$ rises to +1V the I-transistor will conduct and when $V_{control}$ drops to -23.8V, the K-transistor will conduct.

Both conduction-periods are short. The change of the tank-circuit-inductance is controlled by these two signals. The six blocks L to R (all of them flip-flops) keep track of how far the 3-step cycle has progressed. The blocks S and T insure the desired initial position of the six flip-flops, and their resetting when the 3-step cycle is over.

The last blocks perform the two actual short-circuit operations whereby the tank-inductance is changed.

In the "Third Bimonthly Report" the blocks L, N and Q were shown in Figure 19, the blocks M, P and R were shown in Figure 20, and the eight blocks from L to T were shown schematically in Figure 21.

When power is initially applied to the transmitter, the connection between the blocks S and T will have a voltage of about -20V. This low voltage insures that the lowermost transistor (the control-transistor) in the left leg of the flip-flop that constitutes block L is reverse biased. Consequently the flip-flop is monostable with the right leg conducting.

The uppermost transistor in the right leg of block M controls whether the M-flip-flop is mono-or bistable. The transistor's emitter-voltage is independent of which leg is conducting. With the left leg of the L-flip-flop non-conducting the M-control-transistor is reverse-biased, and the M-flip-flop is monostable with the left leg conducting.

In the same way it may be shown that the other 4 flip-flops have to be monostable when voltage is applied to the transmitter; L, N and Q with the right leg conducting, M, P and R with the left leg conducting.

A consequence of this is that the S-transistor is reverse-biased. The T-capacitor will therefore charge to a voltage high enough to forward-bias the L-control-transistor. L is now bistable, but the right leg is still conducting and the five other flip-flops are still monostable. At the moment $V_{control}$ has, for the first time, reached +1 Volt, the collector-voltage of the I-transistor jumps from about -22 volts up to ground-voltage (and subsequently returns to -22 Volts). This impulse applied to L, N and Q. L is bistable and flips, whereas the monostable N and Q are not actuated.

With the left leg of L conducting, the M-control-transistor is forward biased and M is bistable. The left leg of M is still the conducting leg. When $V_{control}$ has dropped to -23.8 Volts, the collector-voltage of the K-transistor jumps from about ground-voltage to -22 Volts, returning subsequently to ground voltage. This impulse is applied to M, P and R. M is bistable and flips, whereas the monostable P and R are not actuated.

With the right leg of M conducting, the N-control-transistor is forward biased and N is bistable. The right leg of N is still the conducting leg.

When the right leg of M became conducting (and it remains so until resetting takes place), the voltage of the connection from M to U was raised enough to saturate the U-transistor. The two W-transistors will start to conduct as A.C. switches as they are being saturated. Thus the lower part of the X-coil (the value of which was L_1) is shorted out. The remaining inductance has the value L_2.

$V_{control}$ does now sweep through its range -23.8V to +1V with a new inductance. ($V_{control}$ continuously changes the value of the two voltage-sensitive capacitors in block X.)

The second time $V_{control}$ reaches +1V, a signal from I is applied to L, N and Q. The actuated L remains actuated (left leg conducting). The bistable N is actuated (and the P-control-transistor is forward-biased, making P bistable). The monostable flip-flop Q still has the right leg conducting, and H collector goes to ground.

The A-capacitor discharges. For $V_{control}$ = 23.8V a signal from K makes P actuate. (M and R are unaffected.)

As the conducting leg of P is now the right one, the Q control transistor is forward biased, and the Z transistor is saturated. Thus the two V transistors are saturated, meaning that a larger portion of the X-coil is shorted out. The remaining inductance has the value L_3. At the end of the third sweep of $V_{control}$, Q is actuated, and when $V_{control}$ again has dropped to -23.8 Volt, R is actuated.

When R is actuated, the S transistor starts to conduct. The connection between S and T is pulled down to -20.5 Volts. (There is a 2 Volt drop across the four S-diodes). The connection between T and L drops enough in voltage to reverse-bias the $L_{control}$ transistor. L resets (the right leg is again conducting) making the 5 other flip-flops reset. When R resets (left leg conducting), the S transistor is again reverse-biased. The operation is fast enough to insure that the capacitor does not discharge appreciably. The $L_{control}$ transistor is, therefore, immediately forward-biased, and the L flip-flop again bistable (with the right leg conducting) followed by the 5 monostable flip-flops.

As a consequence of the resetting, the U and Z transistors are reverse-biased, thus breaking the two short-circuits. (Both A.C. switches Y and W are non-conducting.) The X-coil has again the inductance L_1, and the 3 step cycle has been completed. The circuit will automatically start on a new 3 step cycle.

In case this repetition should not be desired, the blocks R and S would be omitted and the voltage swing of the connection from Q to R would be used to keep the H connection to the modulator at ground-voltage. In this way it could be insured that $V_{control}$ only makes 3 cycles, after which the A-capacitor stands by in the discharged state.

Lock-on and consequent tuning of the transmitter is accomplished by an output (R.F.) driving the switch transistor (lowermost in Figure 3 of the "First Bimonthly Report") into conduction. This then causes the balanced modulator to balance, as charge leaks off capacitor-A. This switch starts to open and more charge is applied to the capacitor-A. Equilibrium will establish itself and the charge will remain constant thereby keeping the transmitter stages tuned by application of the $V_{control}$ voltage to the X capacitors. This state may be realized in any of the 3 cycles previously described. The cycling or searching is of course immediately stopped as soon as a tuning point is reached. To start the cycling again the crystal must be removed from the transmitter.

IV Conclusions

A solution has been found to the impedance matching problem which, on paper, appears to be practical. It has not been built because the values of the variable elements of the pi network cannot be specified until the characteristics of the output transistor are known. A meeting with the customer disclosed that it is probable that information concerning a transistor, coming close to meeting the output requirements, will be available in the not too distant future. In order to satisfy the original specifications for the antenna impedance matching system, a three-variable pi network is necessary. The need for the third variable element introduces appreciable additional complexity. In an ultimate version, where size is of paramount importance, the customer may consider that the small decrease in versatility which results with a simpler two-variable network may not justify the use of the more complex system.

The study of two-variable systems has resulted in the production of data which show the ratio of the maximum-to-minimum values of the elements necessary to meet a variety of conditions. Calculations have been made for matching to any antenna impedance falling within an area defined by the superposition of a spiral on the original impedance rectangle. The spiral represents the impedance of a certain type of antenna as a function of frequency. The resultant area is similar to that originally specified but with the low resistance, high reactance corners cut-off. These calculations have been made for a number of different transistor load impedances.

The automatic transmitter tuning circuitry has been constructed and successfully demonstrated. Now that an operative system is available, modifications have become apparent which would result in a considerable simplification. In the interests of reliability and small size, the tuning system is being reviewed critically to determine where improvements can be made.

V Future Plans

In the area of antenna impedance matching, it will be necessary to consult with the customer concerning which of the various alternative courses of action should be followed. The situation is made more complex by the fact that a transistor suitable for use in the output stage is not presently available. Although it is not anticipated that a great amount of time will elapse before this state of affairs changes, it will be extremely difficult to go ahead with the construction of the matching system until the transistor parameters are definitely known. In particular, the maximum and minimum values required for the elements of the pi matching network are strongly dependent upon the load impedance required by the output transistor.

The customer has requested that the electrical model to be delivered at the conclusion of this program be in a conveniently packaged form of small physical size. It is understood, however, that the equipment will not be constructed in a truly miniaturized form, with maximum component density. As a design goal, throughout this program, solutions to the various circuit problems which do not appear to be compatible with the ultimate size limitations, have been rejected.

However, construction of the equipment in a small rather than a highly miniaturized form will still require the fabrication of many special components such as variable capacitors and inductors. These components will be special in the sense that they are not available in small physical sizes with appropriate electrical characteristics. On the other hand, the electrical specifications are not of such a nature as to preclude miniaturized construction.

VI Identification of Key Technical Personnel

There have been no additions to the personnel during the present reporting period.